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Title:

SYSTEM AND METHOD FOR PROVIDING A LOSSLESS
AND DISPERSION-FREE TRANSMISSION LINE

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SYSTEM AND METHOD FOR PROVIDING A LOSSLESS
AND DISPERSION-FREE TRANSMISSION LINE

CROSS-REFERENCE TO RELATED APPLICATIONS

[0001]   The present application is related to concurrently filed, co-pending, and commonly assigned U.S. Patent Application No. XX/XXX,XXX, Attorney Docket No. 10031009-1, entitled "SYSTEM AND METHOD FOR PROVIDING DISTRIBUTED AMPLIFICATION"; and U.S. Patent Application No. XX/XXX,XXX, Attorney Docket No. 10031011-1, entitled "SYSTEM AND METHOD FOR PROVIDING A DELAY LINE AND/OR FINITE IMPULSE RESPONSE FILTERS USING A LOSSLESS AND DISPERSION-FREE TRANSMISSION LINE", the disclosures of which are hereby incorporated herein by reference.

TECHNICAL FIELD

[0002]   This invention relates to transmission lines and more specifically to systems and methods for providing a lossless and dispersion-free transmission line.

BACKGROUND OF THE INVENTION

[0003]   At high frequencies, for example, at or above 1GHz the propagation of electrical signals along conducting wires is hampered by the effect of losses. Such losses typically cause attenuation as well as dispersion of the signals. If the signal represents a stream of digital data, dispersion causes smoothing of temporal edges, limiting the rate at which digital symbols can be transmitted without intersymbol interference. Attenuation also makes it difficult to identify digital symbols.

[0004]   It has been found useful to use a transmission line for a variety of purposes, one such purpose being as a delay line. However, attenuation and/or dispersion of the signal at high frequencies, limits the usefulness of transmission lines as delay line elements.

[0005]   One system for controlling attenuation and/or dispersion in a primary conductor is by use of an auxiliary conductor inductively coupled to the primary conductor. The auxiliary conductor is driven by the primary conductor through an active shunt network

distributed along the transmission line. In a variation of this system, two pairs of conductors including a first and second primary conductor and a first and second auxiliary conductor can be operated in differential mode. As will be seen later, the distributed active shunt network can be particularly simple in differential mode.

BRIEF SUMMARY OF THE INVENTION

[0006] A lossless (or low loss) transmission line can be constructed using an auxiliary conductor inductively coupled to the primary conductor. The auxiliary conductor is driven by the primary conductor through an active shunt network distributed along the transmission line. The auxiliary conductor is placed close enough to the primary conductor so that the two conductors are inductively coupled (i.e. have a substantial amount of mutual inductance compared to their self-inductance). In a variation of this system, two pairs of conductors including a first and second primary conductor and a first and second auxiliary conductor can be operated in differential mode.

[0007] In one embodiment, a combination of conductance and transconductance are used to cancel losses and control dispersion in the transmission line for high frequency signal transmission. The signal is not assumed to be binary in amplitude, and the transmission line can operate on analog as well as digital signals. In such an embodiment, transconductance is achieved in a differential transmission line by inducing a signal from each transmission line into closely coupled parallel lines, adding active elements between each of the coupled lines to a common ground plane and influencing the current through each active element by the signal on the opposite transmission line.

[0008] In another embodiment the transmission line provides gain while remaining dispersion-free. The total gain grows exponentially with line length and there is no fundamental limit to the length over which the transmission line will provide gain.

[0009] In another embodiment the bi-directional nature of the transmission line enables the implementation of active resonant line segments for use as on-chip frequency references. Thus, an oscillator can be constructed without the use of crystals or other control devices.

[0010]   In still another embodiment, the transmission line can be used as a delay line, for example, in a finite impulse response (FIR) filter.

[0011]   The foregoing has outlined rather broadly the features and technical advantages of the present invention in order that the detailed description of the invention that follows may be better understood. Additional features and advantages of the invention will be described hereinafter which form the subject of the claims of the invention. It should be appreciated that the conception and specific embodiment disclosed may be readily utilized as a basis for modifying or designing other structures for carrying out the same purposes of the present invention. It should also be realized that such equivalent constructions do not depart from the invention as set forth in the appended claims. The novel features which are believed to be characteristic of the invention, both as to its organization and method of operation, together with advantages will be better understood from the following description when considered in connection with the accompanying figures. It is to be expressly understood, however, that each of the figures is provided for the purpose of illustration and description only and is not intended as a definition of the limits of the present invention.

BRIEF DESCRIPTION OF THE DRAWINGS

[0012]   FIGURE 1 shows a schematic representation of one embodiment;

[0013]   FIGURE 2 shows a single slice of the structure of FIGURE 1;

[0014]   FIGURES 3A, 3B, 3C and 3D depict embodiments of lumped implementations of differential and non-differential circuits;

[0015]   FIGURE 4 shows a termination network;

[0016]   FIGURES 5, 6, and 7 show graphs of a test system;

[0017]   FIGURE 8 shows one modification to achieve network gain;

[0018]   FIGURE 9 shows a generalized shunt network;

[0019]   FIGURE 10 shows a representative alternative circuit;

[0020]   FIGURE 11 is an approximation of a transmission line by a ladder network

of inductors and capacitors;

[0021] FIGURES 12A and 12B are illustrated circuit architectures implementing a finite impulse response filter; and

[0022] FIGURE 13 is an equivalent circuit model of an active coupled line.

DETAILED DESCRIPTION

[0023] Before beginning the description, it should be noted that envisioned applications include transmission of critical high-frequency a.c. signals within large chips, or over long distance transmission lines with the concepts taught herein being used in repeaters to boost and control signal dispersion. In addition, accurate delay lines, on-chip oscillators and frequency references, high-speed output drivers and distributed electrostatic discharge (ESD) protection structures, finite impulse response filter and other circuit elements could also be designed around the concepts discussed herein. Amplifier array chips based on these concepts could be inserted in series with long printed-circuit board (PCB) traces in order to split their length and thereby boost the bandwidth of such traces.

[0024] FIGURE 1 shows a schematic representation of one embodiment, in which a symmetrical distributed structure is represented. The distributed structure shown in FIGURE 1 includes differential transmission lines such that transmission line 11^+ carries the exact opposite signal from transmission line 11^-. Coupled to, but not electrically connected with, each transmission line is an auxiliary conductor, such as conductor 12^+ and conductor 12^-. Conductor 13 is the common return path ground. The cross-section structure is essentially invariant along the direction of the x axis. Each conductor 12^+ and 12^- is connected to common ground 13 by a number of conductance and transconductance elements spaced along the transmission line corresponding to points $11a^+$ to $11n^+$ on conductor 11^+ and points $11a^-$ to $11n^-$ on conductor 11^-. In the embodiment shown, transconductance is achieved by controlling a current device, such as current device $14aGm^+$ from the differential "opposite" transmission line. Thus, the transconductance elements associated with transmission line 11^+ are controlled by the signals at the respective point on opposite transmission line 11^-. In FIGURE 1, x is the direction of a.c. transmission and, as will be discussed, can be bi-directional.

[0025] For better clarity, FIGURE 2 shows a schematic of a single slice 20 of the structure shown in FIGURE 1. Conductors 11 and 12 (for convenience, the notation conductor 11 means both conductors 11^+ and 11^- and similarly, conductor 12 means conductors 12^+ and 12^-) are close enough to each other such that the capacitance between them is not much smaller than is the capacitance between conductors 11 and 13 or between conductors 12 and 13. The spacing between the positive and the negative side of the differential transmission pair is not critical. Optimally, the positive and negative conductors should be far apart enough so that the capacitance between them is smaller than the capacitance between them and conductor 13. A factor of 3 smaller would be ample. For example, the spacing of the conductors should be such that the capacitance between conductors 11^+ and 11^- is less than 60pF/m while, as shown in Table 2, the capacitance (C10') between conductors 11 and 13 (ground) is, for example, 173pF/m. The capacitance between conductors 12^+ and 12^- is less than 7pF/m, while the capacitance (C20') between conductors 12 and 13 is, for example, 21.2pF/m. These examples assume the other capacitances have the values shown in Table 2, discussed hereinafter.

[0026] Ideally, active shunt network 21 would be truly distributed along the length of the transmission line. In this case, conductance $14aG^+$ would be a made of continuous resistive material, whereas $14aGm^+$ would be a single, very wide transistor. However, in common integrated circuit technologies, it would be difficult to cross-connect the control electrodes between the two sides of the differential structure in a truly distributed implementation. This is so because lines $14a^+$ and $14a^-$ (as well as lines $14b^+$ and $14b^-$; $14c^+$ and $14c^-$, etc.) each would be continuous and thus physically unable to cross. Instead, a good approximation of the distributed shunt network can be obtained by lumped shunt circuits placed at regular intervals along the transmission lines, as shown in FIGURE 1.

[0027] Each pair of coupled transmission lines is characterized by a set of capacitances and inductances per unit length, as well as the series resistance per unit length of each conductor. These parameters are listed and described in Table 1. As will be discussed, wave propagation in conductors 11^+ and 11^- is lossless and dispersion-free in differential mode if the shunt element values are chosen as follows (the prime mark is used to indicate per unit length):

$$G2' = \frac{R1' \cdot C12'}{M12'}$$

(1)

$$Gm2' = R1' \frac{C10' + C12'}{M12'}$$

(2)

Table 1   Transmission line parameters

Parameter	Description
C10'	capacitance per unit length between conductors 11 and 13
C20'	capacitance per unit length between conductors 12 and 13
C12'	capacitance per unit length between conductors 11 and 12
L1'	self-inductance per unit length of conductor 11
L2'	self-inductance per unit length of conductor 12
M12'	mutual inductance per unit length between conductors 11 and 12
R1'	series resistance per unit length of conductor 11
R2'	series resistance per unit length of conductor 12

[0028]   Lossless and dispersion-free propagation is available over the whole frequency range where the above conditions can be satisfied. It should be noted that a constant resistance per unit length R1' is a reasonable approximation in integrated circuits, where conductors are typically thin compared to skin depth, even at frequencies of several gigahertz.

[0029]   A single-ended (non-differential) lossless transmission line could be implemented using a negative distributed transconductance Gm2'. Unfortunately, such an element is not available, so the differential structure with cross-coupled control electrodes 14[+] and 14[-] effectively emulates a transconductance Gm2' for the differential component of the wave in conductors 11[+] and 11[-]. Any common-mode component will be affected by losses and decay as the wave travels along the transmission line.

[0030]   FIGURE 3A depicts one embodiment 30 of a lumped active shunt network. An instance of this circuit or its equivalent is placed at regular intervals Δx along the transmission line. In FIGURE 1, these intervals would be at locations 11a, 11b, 11c to 11n and would be the same on both differential lines (+ and -). Lumped conductance G2 (14aG[+]) has a value $G2' \cdot \Delta x$. Similarly, transistors N1 (301) and N2 (302) should have a transconductance Gm2 equal to $Gm2' \cdot \Delta x$. This transconductance can be adjusted, for example, by using input terminal bias (304) and transistor N3 (303). Instead of using MOSFETs, as shown, this circuit

could as well be implemented using other types of transistors or even other circuit elements.

[0031] In FIGURE 3A, conductors 11^+ and 11^- are shown on top of conductors 12^+ and 12^- respectively. This configuration is advantageous in integrated circuit technologies offering a thicker (or wider) metal layer at the top. The resistance per unit length $R1'$ of conductors 11^+ and 11^- determines how lossy the passive transmission line would be, hence how much power must be spent by the active network in order to compensate for those losses. For this reason, it is advantageous to allocate the largest possible cross-section to conductors 11^+ and 11^-. Conversely, there is no particular reason why conductors 12^+ and 12^- should have a low resistance per unit length, therefore it is acceptable to keep them thin and narrow.

[0032] It is well known that the association of a transconductance Gm with a conductance G constitutes an elementary amplifier with a voltage gain A given by

$$A = -\frac{Gm}{G} \qquad (3)$$

and an output impedance $Zout$ given by

$$Zout = \frac{1}{G} \qquad (4)$$

[0033] Therefore, the active shunt network shown in FIGURE 3A can be replaced by other circuits having the function of an amplifier with the appropriate gain and output impedance as shown in FIGURE 3B. The distributed nature of the ideal active shunt network can be approximated by placing a lumped amplifier such as 310, 311 at regular intervals Δx along the transmission line. The interval Δx must be much smaller than the shortest wavelength of interest for the application. This is also true of the embodiment shown in FIGURE 3A. To achieve conditions of lossless propagation, each amplifier must have a voltage gain

$$A = -\frac{Gm2'}{G2'} = -\left(1 + \frac{C10'}{C12'}\right) \qquad (5)$$

and an output impedance

$$Zout = \frac{1}{\Delta x \cdot G2'} = \frac{M12'}{\Delta x \cdot R1' \cdot C12'} \qquad (6)$$

[0034] In this generalized framework, it can be seen that a single-ended (i.e. non-differential) version of the invention can be implemented as well. The schematic of a cross-

section of a shunt network for the single-ended transmission line is shown in FIGURE 3C. The main difference is that the amplifier constituting the active shunt network must have a positive gain in this case (non-inverting amplifier). Unfortunately, a positive gain can be obtained in practice only by cascading two amplification stages with negative gains, which makes the single-ended version somewhat less attractive than the differential version. A reasonable implementation of a single-ended shunt network as shown in FIGURE 3D, which is a slightly modified version of the circuit shown in FIGURE 3A, yielding a non-inverting amplification stage. The input terminal *ref* must be set to a constant voltage approximately equal to the DC component of the signal traveling in conductor 11. If one thinks in the framework of amplifiers instead of transconductances, then the benefit of the differential structure is that the amplifiers can be made inverting (single stage) instead of non-inverting by cross-connecting their inputs, whereby a single stage amplifier can be used for each side.

[0035]   To avoid reflections at the extremities, each pair of coupled transmission lines (11^+, 12^+ and 11^-, 12^-) is terminated by the lumped network shown in FIGURE 4. At the driven end, a voltage source , such as source 41, can be inserted ahead of conductor 13 and resistance $R0$ can be partially or totally absorbed in the output impedance of the voltage source, if desired.

[0036]   As will be discussed, reflections will be totally cancelled if the element values are the following:

$$R0 = v_{ph}M12' \tag{7}$$

$$R1 = v_{ph}\left(L1'-M12'\right) \tag{8}$$

$$C1 = \frac{1}{v_{ph}R1'} \tag{9}$$

[0037]   Transmission line parameters defined above in Table 1 apply again in this case. The phase velocity v_{ph} is equal to the speed of light in the dielectric material surrounding the transmission line conductors. It depends only on the dielectric constant of this material.

[0038]   The nature and value of impedance Z2 terminating conductor 12 does not affect propagation conditions in conductor 11. The range of possible choices includes the short circuit case (Z2=0), as well as the open-circuit case (conductor 12 left unconnected).

[0039] A 48mm long transmission line (which could, for example, be used for a delay line, perhaps in a finite impulse response filter) has been simulated using a circuit simulator. This corresponds to a nominal delay of about 320ps. The line was modeled by 4800 cascaded segments consisting of lumped capacitors, inductors and resistors. Numerical parameters for this line are listed in Table 2. They have been calculated using a finite-element approach for a 3.8μm wide line using the top four metal layers of a 0.13μm CMOS process.

Table 2

Parameter	Value
C10'	173pF/m
C20'	21.2pF/m
C12'	97.7pF/m
L1'	228nH/m
L2'	518nH/m
M12'	187nH/m
R1'	2.73KΩ/m
R2'	109KΩ/m
G2'	1.43A/Vm
Gm2'	3.95A/Vm

[0040] In a first simulation, the active shunt networks (such as networks 21^+ and 21^- in FIGURE 2) were left out. A 1.25GHz square wave with a peak-to-peak amplitude of 200mV (differential) was applied to the input of the termination network. The voltage at nine different taps (51-59) at 6mm intervals is shown in FIGURE 5. As can be seen, edge amplitude decays rapidly with distance from the driving point.

[0041] In a second simulation, shown in FIGURE 6, the active shunt networks are added. Transconductances are modeled by voltage-controlled current sources. Therefore, the shunt network does not capacitively load the line, and the transconductance value is exact at any frequency. The waveforms at the same points (51-59) are shown in FIGURE 6. The result is virtually ideal, free of attenuation and dispersion. This result confirms the expectation from theory that a lossless and dispersion-free transmission line can be achieved within the frequency range where the transconductances can operate properly. Some moderate ringing, which can be seen at edges 61, 62, 63, appears to be an artifact of modeling the transmission line by discrete inductances and capacitances. Such ringing also occurs in Spice simulations of conventional transmission lines when losses are very low and edges very sharp.

[0042] In a third simulation, shown in FIGURE 7, transconductances are

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implemented by actual MOS transistors following the schematic shown in FIGURE 3A. The oscillations in 71, 72 and 73 are again ringing, and are more visible here than in FIGURE 6 because the ringing frequency is lower due to capacitive loading. Reflections can be seen in the bumps and dips occurring after the first 320ps of the simulation. The waveforms look almost perfect for the first 320ps because the transmission line was initially zero at all points. After traveling across the whole line (320ps), the wave hits the termination and is partially reflected back because the termination does not perfectly match the transmission line characteristic impedance. The reflected wave adds to the forward wave which keeps coming from the source. Within ellipses 72 and 73, one can quite distinctly see that the perturbation is also a square wave which supports the view that the distortions are due to the addition of the reflected wave.

[0043]   Some amount of dispersion can also be observed on traces corresponding to taps remote from the driving point. Comparing FIGURE 7 to FIGURE 6 and looking at the first 320ps of simulation, it can be seen that the slope of the signal becomes gradually less steep as the wave propagates along the transmission line. In a perfectly non-dispersive situation, the waveform would remain identical during propagation. The likely reason for this imperfection is that the transconductance of MOSFETs decreases above some corner frequency due to channel transit time effects. Above this frequency, the conditions for lossless and dispersion-free propagation may not be able to be fully met.

[0044]   Instead of only compensating for losses, it is possible to obtain gain from a slightly different version. FIGURE 8 shows one possible modification to achieve network gain. This modification consists of adding distributed conductance 81^+, 81^- between conductors 11^+, 11^- and conductor 13, and modifying the value of transconductance $Gm2'$. In order to achieve a specific gain α (measured in Neper per unit length), element values should be chosen as follows:

$$G1' = \frac{\alpha^2}{R1'} \qquad (10)$$

$$G2' = \frac{R1' \cdot C12'}{M12'} \qquad (11)$$

$$Gm2' = R1' \frac{C10' + C12'}{M12'} + \frac{2\alpha}{v_{ph} M12'} - \frac{L1'}{M12'} \frac{\alpha^2}{R1'} \qquad (12)$$

[0045]   Unlike conventional distributed amplifiers, there is no fundamental limit to

the length over which this line will provide gain. However, reflections at the terminations would cause the amplifier to oscillate if the total gain multiplied by the reflection coefficient exceeded unity. Therefore, the maximum achievable gain is effectively limited by the accuracy with which reflections can be cancelled (or the reflections contained) at the terminations.

[0046] FIGURE 9 shows generalized shunt network 90 as a theoretical model for identifying other circuits which can control lossless transmission. This shunt network is made of conductances and transconductances between conductors 11^+, 12^+, and 13. In FIGURE 9, V1 is the voltage between conductors 11^+ and 13, and V2 is the voltage between conductor 12^+ and 13. As will be discussed, wave propagation on conductor 11^+ will be dispersion-free and present a gain per unit length α if the element values meet the following conditions:

$$Gm1'-G12'-Gm12'= 0 \tag{13}$$

$$G2'+G12'+Gm12'=\frac{R1'\cdot C12'}{M12'} \tag{14}$$

$$-Gm2'+G12'+Gm21'\geq \frac{L1'}{M12'}\frac{\alpha^2}{R1'}+R1'\cdot(C10'+C12') \tag{15}$$

$$G1'+G12'+Gm21'=\frac{M12'}{L1'}\left(-Gm2'+G12'+Gm21'-R1'\cdot(C10'+C12')\right) \tag{16}$$

[0047] There are seven free parameters and only four constraints, therefore there are potentially many choices. The solution described with respect to FIGURE 1 corresponds to the case where Gm1', G12', Gm12', Gm21' and G1' are all set to zero. This is believed to be the simplest circuit meeting these constraints, but other solutions exist.

[0048] It is not possible to represent all possible circuits meeting these constraints in a single schematic because the connectivity of the network depends on the signs of element values. Negative element values can be implemented, for example, by cross-connecting the element between the two sides of a differential structure, as shown in FIGURES 1 and 2, whereas positive parameters correspond to elements connected solely to one side of the structure. It would be a tedious job to draw all possible solutions. Instead, just a single alternative solution is shown in FIGURE 10 for the purpose of illustrating how diverse other solutions can become. This circuit corresponds to G12'=0, Gm12'=0 and Gm2'=0, which leads to a positive Gm21', a positive G2' and a negative G1'. This circuit has practical drawbacks compared to the preferred implementation, but in principle it would also work.

[0049] The distributed structure described in FIGURE 1 can be approximated by a circuit made entirely of lumped elements. This lumped approximation is useful if the purpose of the circuit is to delay a signal by a specific amount, rather than transmitting the signal from one point to another. In this case, the coupled transmission lines are replaced by a ladder network of coupled lumped inductors and capacitors. A discrete active shunt network circuit is added at each node of the ladder network in order to cancel resistive losses in the inductors.

[0050] A schematic of such an embodiment is shown in FIGURE 11. The dashed arrows indicate that the controlled current sources $Gm2^+$ are controlled by the voltage on conductor 11- on the opposite side of the differential structure. Two stages of a single-ended network are shown, but the network may have any number of such stages in cascade. Only one side (the positive side) of the differential structure is shown. The second (negative) side is identical. Only intentional circuit elements are shown. The inductors unavoidably have some parasitic resistance in series which causes losses. The active shunt networks consisting of a conductance G2 and a controlled current source Gm2 cancel these losses. In practice, the active shunt networks can be implemented by the same circuit as shown in FIGURE 3.

[0051] It can be shown that network 1100 constitutes a delay line. The signal applied to conductor 11 at one end of the line propagates at a constant velocity along the network. Each stage introduces a delay Δt given by

$$\Delta t = \sqrt{L1 \cdot \left(C10 + \frac{C12 \cdot C20}{C12 + C20} \right)} \qquad (17)$$

[0052] Therefore, each stage approximates a length $v_{ph} \cdot \Delta t$ of distributed transmission line, where v_{ph} is the velocity of light in the medium under consideration.

[0053] In order to achieve lossless and dispersion-free propagation, element values shown in FIGURE 11 must be related as specified in the following equations:

$$G2 = \frac{R1 \cdot C12}{M12} \qquad (18)$$

$$Gm2 = R1 \frac{C10 + C12}{M12} \qquad (19)$$

$$M12 = \frac{C12}{C12 + C20} L1 \qquad (20)$$

[0054] A number of possible applications of this concept exist. One such is in the transmission of high-speed signals across a large chip. Another is an amplifier. Two other applications will be described with respect to FIGURES 12A and 12B.

[0055] A finite impulse-response filter computes a discrete weighted sum of delayed copies of its input signal:

$$Y(t) = \sum_{k=1}^{N} W_k \cdot x(t - \Delta t_k) \tag{21}$$

[0056] The operation performed by the finite impulse response filter depends on the values of coefficients W_k, known as *tap weights*. The delays Δt_k are typically integer multiples of some unit delay. Two possible analog implementations of a finite impulse response filter are illustrated in FIGURES 12A and 12B.

[0057] FIGURE 12A shows the input signal 1201 applied to delay line 1202. Taps along this line provide delayed copies of the input signal. The voltage at each tap is applied to a transconductance amplifier such as 1203-1 delivering a current proportional to the voltage. The ratio Gmk between current and voltage determines the tap weight. The currents from all transconductance amplifiers are added together on a global node *output* 1204 loaded by resistor R_{load}. The voltage on this global node is the output signal of the filter. This circuit works in principle, but it may be difficult in practice to reach very high speeds because the parasitic capacitance of the global output node tends to be large.

[0058] A slightly modified architecture is shown in FIGURE 12B. In this variation, delay elements 1205 are added between the outputs of the transconductance amplifiers, whereby the delays on the input side are correspondingly reduced in order to maintain the same total delay as the signal travels from the input to the output through a given tap. This circuit operates in much the same way as does the circuit shown in FIGURE 12A, but can achieve much higher bandwidth because the tap outputs are aggregated over a delay line. The reason is that the parasitic capacitance unavoidably present at the output of each transconductance amplifier becomes part of the delay line capacitance in this case. An implementation of this architecture using passive (i.e. lossy) inductor/capacitor ladders as delay lines has been published recently (see e.g., Wu, et al., "Differential 4-Tap and 7-Tap Transverse Filters in SiGe for 10Gb/s Multimode Fiber Optic Link Equalization", *IEEE Int. Solid-State Circ. Conf.*, San Francisco, Feb. 2003, hereinafter referred to as *Wu*). The *Wu* filters could be implemented using the

techniques shown in FIGURES 12A and 12B.

[0059]   Both architectures could be implemented either in the form of a distributed transmission line, or in the form of a lumped approximation.

[0060]   As discussed above, the transmission line concepts can be used as an amplifier. If both ends of the transmission line are terminated as described with respect to FIGURE 4, then a signal applied to one end will travel only one way across the line, and will be fully absorbed by the termination. If the termination does not perfectly match the characteristic impedance of the line, then at least a fraction of the signal will be reflected back and amplified again on the return to the source. Again, if the source impedance does not match the characteristic impedance of the line, a fraction of the signal is reflected back into the line.

[0061]   If the total gain of the amplifying line is A, and if a fraction α of the signal is reflected back at each termination, then the line will become unstable if the product $\alpha \cdot A$ exceeds unity. In this case, an oscillation will build up from noise and the line can be used as an oscillator. The fundamental oscillation frequency f_0 will be

$$f_0 = \frac{v_{ph}}{2L} \qquad (22)$$

where v_{ph} is the speed of light in the considered medium and L is the length of the line. The oscillator output will generally also include harmonics of this frequency.

[0062]   An economical way to produce such an oscillation would be to leave both ends of the line open (no termination), whereby α would approach unity. In this case, gain just slightly larger than unity should suffice to produce sustained oscillations.

[0063]   In an integrated circuit, both v_{ph} and L can be accurately controlled and stable over time and environmental conditions, therefore it should be possible to use such an oscillator as an on-chip frequency reference.

[0064]   Wave propagation characteristics of a pair of coupled lines with a distributed active shunt network will now be derived. FIGURE 13 (shown on the page with FIGURE 4) shows equivalent circuit 1300 of a short segment of the structure under consideration. Distributed capacitances C10', C20' and C12', inductances L1', L2', M12' and series resistances R1' and R2' are inherent to the transmission line conductors. Shunt conductances G1', G2' and transconductance Gm2' are added in an attempt to make the line

dispersion-free and lossless. The ground conductor 13 carrying the return currents is not explicitly shown.

[0065] It should be noted that the set of transmission line parameters is redundant if propagation is purely transverse electro-magnetic (TEM). Inductance parameters L1', L2' and M12' are related to capacitance parameters C10', C20' and C12'. It can be shown that the relationship between them is:

$$L1' = \frac{C12'+C20'}{C10'C20'+C10'C12'+C20'C12'} \cdot \frac{\varepsilon_r}{c_0^2} \tag{23}$$

$$L2' = \frac{C12'+C10'}{C10'C20'+C10'C12'+C20'C12'} \cdot \frac{\varepsilon_r}{c_0^2} \tag{24}$$

$$M12' = \frac{C12'}{C10'C20'+C10'C12'+C20'C12'} \cdot \frac{\varepsilon_r}{c_0^2} \tag{25}$$

[0066] In these equations, c_0 is the speed of light in vacuum and ε_r is the dielectric constant of the medium surrounding the conductors.

[0067] The voltage gradient at a given point of the transmission line is related to the currents at this point as follows:

$$\begin{cases} \frac{\partial V_1}{\partial x} = -(R1'+sL1') \cdot I_1 - sM12' \cdot I_2 \\ \frac{\partial V_2}{\partial x} = -sM12' \cdot I_1 - (R2'+sL2') \cdot I_2 \end{cases} \tag{26}$$

[0068] Similarly, the current gradient can be written:

$$\begin{cases} \frac{\partial I_1}{\partial x} = -(G1'+s(C10'+C12')) \cdot V_1 + sC12' \cdot V_2 \\ \frac{\partial I_2}{\partial x} = (-Gm2'+sC12') \cdot V_1 - (G2'+s(C20'+C12')) \cdot V_2 \end{cases} \tag{27}$$

[0069] It is helpful to rewrite these equations in matrix form:

$$\begin{cases} \frac{\partial}{\partial x}\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = -\mathbf{Z} \cdot \begin{pmatrix} I_1 \\ I_2 \end{pmatrix} \\ \frac{\partial}{\partial x}\begin{pmatrix} I_1 \\ I_2 \end{pmatrix} = -\mathbf{Y} \cdot \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \end{cases} \tag{28}$$

where

$$Z = \begin{pmatrix} R1'+sL1' & sM12' \\ sM12' & R2'+sL2' \end{pmatrix} \qquad (29)$$

$$Y = \begin{pmatrix} G1'+s\cdot(C10'+C12') & -sC12' \\ Gm2'-sC12' & G2'+s\cdot(C20'+C12') \end{pmatrix} \qquad (30)$$

[0070]   Currents can be eliminated from equation (28) in order to get a second-order differential equation of voltages:

$$\frac{\partial^2}{\partial x^2}\begin{pmatrix} V_1 \\ V_2 \end{pmatrix} = -\gamma^2 \begin{pmatrix} V_1 \\ V_2 \end{pmatrix} \qquad (31)$$

where

$$\gamma^2 = Z \cdot Y \qquad (32)$$

[0071]   In order to achieve dispersion-free wave propagation in conductor 11, the matrix γ^2 must have the following form:

$$\gamma^2 = \begin{pmatrix} \left(-\alpha + \dfrac{s}{v_{ph}}\right)^2 & 0 \\ \gamma^2_{21} & \gamma^2_{22} \end{pmatrix} \qquad (33)$$

the gain exponent α must be real and positive or null. The quantity v_{ph} is the phase velocity of the wave in conductor 11. Assuming that a matrix γ^2 of this form can be achieved, the solution of equation (31), as far as conductor 11 is concerned, can be written:

$$V_1(x) = A_f \cdot \exp(-\gamma_1 x) + A_r \cdot \exp(\gamma_1 x) \qquad (34)$$

[0072]   Parameters A_f and A_r are constants which must be determined using boundary condition at both ends of the transmission line. The propagation exponent is

$$\gamma_1 = -\alpha + \frac{s}{v_{ph}} \qquad (35)$$

[0073]   The form of this propagation exponent implies that the magnitude of the wave traveling across conductor 11 grows exponentially with distance and has linear phase, hence no dispersion.

[0074] It remains to be shown that an appropriate choice of G1', G2' and Gm2' will indeed produce an exponent matrix γ^2 of the desired form. For this purpose, the first row of the matrix product $Z\,Y$ must be written out in detail using equations (29) and (30) and made equal to the first row of matrix γ^2 defined in equation (33). The resulting set of equations must be solved for G1', G2' and Gm2'. This simple but somewhat tedious procedure leads to the result that γ^2 has the required form if the following relations are satisfied:

$$G1' = \frac{\alpha^2}{R1'} \tag{36}$$

$$G2' = \frac{R1'\,C12'}{M12} \tag{37}$$

$$-Gm2' = R1'\frac{C10'+C12'}{M12'} - \frac{2\alpha}{v_{ph}M12'} + \frac{L1'}{M1'}\frac{\alpha^2}{R1'} \tag{38}$$

[0075] As one could have expected, the phase velocity v_{ph} of the wave traveling across conductor 11 turns out to be:

$$v_{ph} = \frac{c_0}{\sqrt{\varepsilon_r}} \tag{39}$$

[0076] Equations (36)-(38) were already introduced above as equations without demonstration.

[0077] If the generalized active shunt network from FIGURE 9 is considered instead of the simpler version of FIGURE 13, matrix Y will be somewhat different, but the same procedure can be applied to find the result presented above.

[0078] The characteristic impedance of a pair of coupled lines cannot generally be expressed by a single scalar, but rather by a matrix Z_c. This matrix is the ratio between the series impedance matrix Z defined in equation (29) and the propagation exponent of the wave traveling in the line.

[0079] In the lossless case, $\alpha = 0$, and the characteristic impedance matrix has the following form:

$$Z_c = \begin{pmatrix} \dfrac{v_{ph}R1'}{s} + v_{ph}L1' & v_{ph}M12' \\ Z_{c21} & Z_{c22} \end{pmatrix} \qquad (40)$$

[0080] The terms Z_{c21} and Z_{c22} can be calculated but the resulting expressions are rather intricate and of little relevance to signal propagation on conductor 11. For this reason, the detailed expressions are not provided here.

[0081] A line of finite length must be terminated by a lumped network characterized by the same impedance matrix as Z_c. The termination network shown in FIGURE 4 meets this constraint. If some finite gain or loss is present ($\alpha \neq 0$), the characteristic impedance matrix becomes more complicated and any termination network that would match the impedance of the coupled line perfectly is not a simple matter. Thus, α should be as small as possible to achieve decent results.

[0082] The disclosed transmission line is characterized by the propagation exponent shown in equation (35), which corresponds to the form required to achieve dispersion-free propagation. Equation (35) is written in the Laplace domain and therefore uses the Laplace variable s which is equal to jw. A negative sign for α is used in equation (35) because the circuit achieves gain. In the more conventional form (positive sign in front of α), a positive value of α means that there are losses, whereas a negative value means that there is gain. Knowing that the invention has gain, it is best to use the opposite convention so that positive values of α mean gain.

[0083] Note that the concepts discussed herein become relevant at frequencies where the length of the transmission line reaches or exceeds about ¼ wavelength. With a signal of 1GHz in a medium where electromagnetic waves propagate at about half the speed of light in vacuum, a wire starts to look like a transmission line if its length exceeds about 37.5mm. Losses become important at this point. The value of 1GHz is reasonable for use inside integrated circuits or on printed circuit boards. If long-distance propagation is considered (on the order of meters or even kilometers), then transmission line theory applies at much lower frequencies, and losses would have to be considered as well.

[0084] Although the present invention and its advantages have been described in detail, it should be understood that various changes, substitutions and alterations can be made herein without departing from the invention as defined by the appended claims. Moreover, the

present application is not intended to be limited to the particular embodiments of the process, machine, manufacture, composition of matter, means, methods and steps described in the specification. As one will readily appreciate from the disclosure, processes, machines, manufacture, compositions of matter, means, methods, or steps, presently existing or later to be developed that perform substantially the same function or achieve substantially the same result as the corresponding embodiments described herein may be utilized. Accordingly, the appended claims are intended to include within their scope such processes, machines, manufacture, compositions of matter, means, methods, or steps.